Losses in Y-Junction Stripline and Microstrip Ferrite Circulators

Robert E. Neidert, *Member, IEEE,* and Purobi M. Phillips, *Member, IEEE*

Abstract— **This paper describes a complete theory to calculate the dielectric, conductor, and magnetic losses of Y-junction stripline and microstrip ferrite circulators in terms of 3-port S-parameters. Calculations using this theory are presented, and compared with existing data for varied circulator parameters over a frequency range of 1–100 GHz.**

I. Introduction

NEW interest is evident in modernization of ferrite component technology. Monolithic microwave circuit technology, which was perfected in the 1980's, has permitted the development of multidecade bandwidth amplifiers, multioctave bandwidth subsystem components, and full systems that have more than an octave bandwidth. The Y-junction circulator is the workhorse of ferrite components. It can be used as a 3-port circulator, 2-port isolator, or a single-pole double-throw switch. For new circulator technology, it would be useful to increase the bandwidth capability (currently approximately an octave), minimize the size (perhaps by using matching circuits other than quarter-wavelength transformers), and create design approaches that are compatible with monolithic circuits. The first step is to have on hand an accurate theoretical analysis tool to calculate intrinsic circulator performance, thereby permitting simultaneous circulator and matching circuit optimization.

A theory to include full dissipation loss calculations has been developed in fundamental ways so that 3-port S-parameters of Y-junction stripline and microstrip circulators can be calculated in the presence of dielectric, conductor, and magnetic losses. A computer program for the analysis of lossless circulators is publicly available [1]. It gives virtually exact agreement with other published calculations for lossless cases. This paper presents the theory involved in the loss calculations, and describes the manner in which the theory has been incorporated into the above computer program. Also, calculations, with loss effects included, are compared with existing data.

Landmark theoretical works on Y-junction circulators include the following:

- Bosma [2], who solves the simplified boundary-value problem of the circulator ferrite disk by using a Green's function approach that relates the axial component of the electric field to the circumferential component of the magnetic field at the perimeter of the disk;
- Fay and Comstock [3], who describe the operation in terms of counter-rotating propagation modes in the ferrite disk;
- Wu and Rosenbaum [4], who predict octave bandwidth microstrip circulator operation theoretically using Bosma's Green's function analysis; and
- Schlömann and Blight [5], who rederive the Bosma form of the Green's function for negative effective permeability ($\mu_{\text{eff}} < 0$).

Some of the critical experimental stripline junction circulator work has been the following:

- Simon [6], who gives the results of tests on a wide range of materials;
- Salay and Peppiatt [7], who clarify the sign of the reactive part of the input impedance; and
- Schlömann and Blight [5] who show very wideband performance over 3:1 bands.

Further background information on ferrite theory and approximate circulator synthesis is compactly provided by Soohoo [8] and Linkhart [9].

II. Stripline and Microstrip Junction Circulators

The field problem for the 3-port symmetrical Y-junction stripline circulator has been addressed by Bosma [2] using the notation in Fig. 1. He developed the form of Green's function relating $\boldsymbol{E}_z$ to $\boldsymbol{H}_\phi$ among all the ports for lossless circulators and for $\mu_{\text{eff}} > 0$. Assuming $e^{j\omega t}$ harmonic time dependence, Bosma's Green's function takes the following form:

$$G(r,\phi;R,\phi') = \frac{jZ_{\text{eff}}J_0(kr)}{2\pi J_0'(x)} - \frac{Z_{\text{eff}}}{\pi} \cdot \sum_{n=1}^{\infty} \frac{\frac{\kappa}{\mu}\frac{nJ_n(x)}{x}\sin n(\phi-\phi') - jJ_n'(x)\cos n(\phi-\phi')}{\{J_n'(x)\}^2 - \left\{\frac{\kappa}{\mu}\frac{nJ_n(x)}{x}\right\}^2} \cdot J_n(kr), \quad (1)$$

where $x = kR$ and $Z_{\text{eff}} = [\mu_0\mu_{\text{eff}}/(\epsilon_0\epsilon_f)]^{1/2}$; ϵ_f being the relative permittivity of the ferrite material, and k the propagation constant, $\omega(\mu_0\epsilon_0\mu_{\text{eff}}\epsilon_f)^{1/2}$.

Schlömann and Blight [5] rederived the Green's function for $\mu_{\text{eff}} < 0$. Most circulators above approximately 2 GHz have relatively small internal bias field, and the operating frequency is far above the ferromagnetic resonance frequency. For this case, the imaginary parts of μ and κ approach zero, the real part of μ approaches unity, and the real part of κ is approximately $\gamma M_s/\omega$ (using rationalized MKS units) as shown in [3, Fig. 5], where γ is the gyromagnetic ratio,

Manuscript received August 3, 1992; revised November 19, 1992.
R. E. Neidert is with the Naval Research Laboratory, Washington, DC 20375-5000. P. M. Phillips is with Science Applications Int'l Corp., McLean, VA 22102.
IEEE Log Number 9209337.

0018–9480/93$03.00 © 1993 IEEE

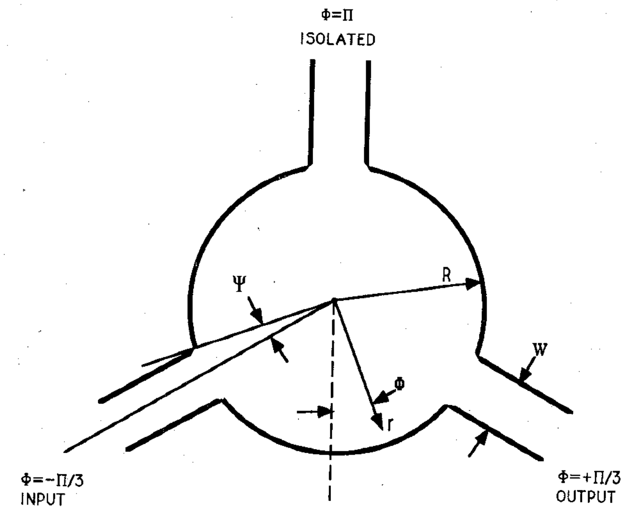


Fig. 1. Circulator coordinate reference.

and M_s is the saturation magnetization. When $|\kappa|$ becomes greater than unity, which happens at frequencies below $f_m = (\gamma/2\pi)M_s$, then $\mu_{\text{eff}} = (\mu^2 - \kappa^2)/\mu$ becomes negative. In order to make calculations possible below f_m, they rederived the Green's function terms for $\mu_{\text{eff}} < 0$. The result is that the terms V_0, V_n, and U_n of the Green's function written in the following form:

$$G(r,\phi;R,\phi') = V_0 + \sum_{n=1}^{\infty}[V_n \cos n(\phi-\phi') + U_n \sin n(\phi - \phi')] \quad (2)$$

change such that each sign is reversed, each Bessel function $J_n(kR)$ becomes a modified Bessel function $I_n(kR)$ of the same order, each Bessel function derivative becomes a modified Bessel function derivative of the same order, and the positive of μ_{eff} is used.

Since no radial current flows from the edge of the center conductor, Bosma suggested the following boundary condition:

$$H_\phi(R,\phi) = \begin{cases} H_a) & -\pi/3 - \Psi < \phi < -\pi/3 + \Psi, \\ H_b) & \pi/3 - \Psi < \phi < \pi/3 + \Psi, \\ H_c) & \pi - \Psi < \phi < \pi + \Psi, \\ zero) & \text{elsewhere}. \end{cases} \quad (3)$$

Following Bosma, and using the condition for cyclic symmetry

$$G(\phi + 2\pi/3; \phi' + 2\pi/3) = G(\phi;\phi'), \quad (4)$$

but *not* using the condition for losslessness

$$G(\phi';\phi) = -G^*(\phi;\phi'), \quad (5)$$

if three quantities C_1, C_2, and C_3 are introduced such that

$$C_1 = \frac{j\pi Z_d}{2Z_{\text{eff}}} + \frac{j\pi}{Z_{\text{eff}}}\Psi G(-\pi/3;-\pi/3)$$
$$C_2 = \frac{j\pi}{Z_{\text{eff}}}\Psi G(-\pi/3;\pi/3)$$
$$C_3 = \frac{j\pi}{Z_{\text{eff}}}\Psi G(-\pi/3;\pi), \quad (6)$$

then, after some algebraic manipulation, one obtains the following expressions for the three coefficients of the scattering matrix:

$$S_{11} = 1 + \frac{\pi Z_d(C_1^2 - C_2C_3)}{jZ_{\text{eff}}(C_1^3 + C_2^3 + C_3^3 - 3C_1C_2C_3)}$$
$$S_{21} = \frac{\pi Z_d(C_2^2 - C_1C_3)}{jZ_{\text{eff}}(C_1^3 + C_2^3 + C_3^3 - 3C_1C_2C_3)}$$
$$S_{31} = \frac{\pi Z_d(C_3^2 - C_1C_2)}{jZ_{\text{eff}}(C_1^3 + C_2^3 + C_3^3 - 3C_1C_2C_3)} \quad (7)$$

where the scattering matrix S is

$$S = \begin{pmatrix} S_{11} & S_{31} & S_{21} \\ S_{21} & S_{11} & S_{31} \\ S_{31} & S_{21} & S_{11} \end{pmatrix}. \quad (8)$$

The expressions for the coefficients S_{11}, S_{21}, and S_{31} are identical to those obtained by Wu and Rosenbaum [4], but the C_1, C_2, and C_3 expressions above, unlike those in [4], are valid for $\mu_{\text{eff}} < 0$ by using the Green's function for that case. There is a small difference between the expressions for C_1, C_2, and C_3 obtained in [4] and those here, in which [4] uses $\frac{\sin^2 n\Psi}{n^2\Psi}$ in the summation of terms in the expression for the Green's function and Ψ is shown here. In the limit $\sum_{n=1}^{\infty} \frac{\sin^2 n\Psi}{n^2\Psi}$ approaches Ψ so that the two are interchangeable if n is reasonably large. Differences between the two are small in calculations using n from 3 to 18. The computer program discussed in this paper is set up to use $\frac{\sin^2 \Psi}{n^2\Psi}$ in the numerical summation as in [4].

Very good agreement has been achieved with Wu and Rosenbaum's lossless calculations using $n = 3$, and with Schlömann and Blight's lossless calculations with negative μ_{eff} below 5 GHz, and with $n = 9$, as shown in Figs. 2 and 3, respectively. The low-frequency extension below f_m is believed to be valid down to about $f_m/2$ as described in [5].

III. Theory of Loss Calculations

A. Dielectric Losses

Modification of the lossless theory to account for dielectric losses is done by introducing the complex dielectric constant.

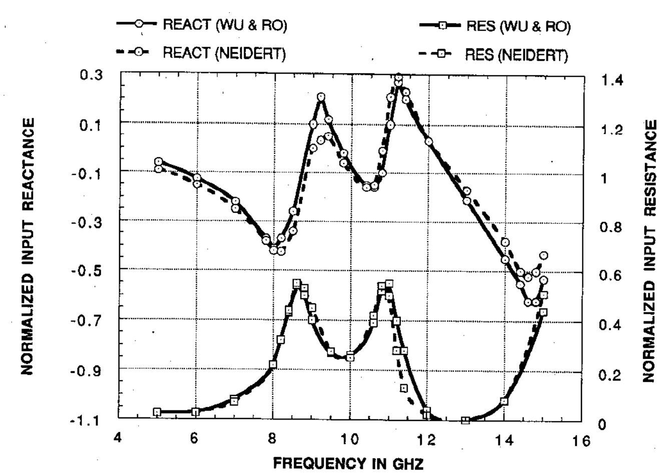


Fig. 2. Comparison to Wu and Rosenbaum.

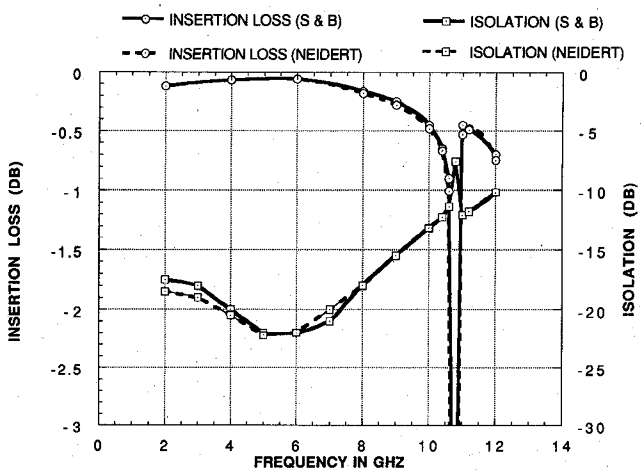


Fig. 3. Comparison to Schlömann and Blight.

It is substituted for the real part used in lossless calculations, and all subordinate calculations are carried through in complex form. The complex dielectric constant is given by $\epsilon = \epsilon' - j\epsilon''$, where $\epsilon'' = \epsilon' \tan\delta$, and $\tan\delta$ is the loss tangent of the ferrite material.

B. Metal Conductor Losses

If the conductor is not perfect, then at the ferrite–metal interface, in addition to the normal component of electric field and tangential component of magnetic field, there exist a small tangential component of electric field and a small normal component of magnetic field. The basic philosophy behind the conductor loss calculation is to find the new resultant electric field $\boldsymbol{E}_N = \boldsymbol{E}_Z + \boldsymbol{E}_{\|}$, noting that $\boldsymbol{E}_{\|}$, like all the other fields in the ferrite according to the original Bosma [2] assumption, has no variation in the vertical direction. Then, following Bosma, it is possible to derive an expression for $\Delta\boldsymbol{E}_N(r,\phi)$ which is related to $H_N(R,\phi')\Delta\phi'$, where $\boldsymbol{H}_N = \boldsymbol{H}_\phi + \boldsymbol{H}_\perp$, and $H_N(R,\phi')$ is the value of $H_N(R,\phi)$ at azimuth ϕ' over a small angle $\Delta\phi'$. The transfer function between $\Delta\boldsymbol{E}_N(r,\phi)$ and $H_N(R,\phi')$ is the new Green's function when conductor loss is present. Once the form of a new Green's function is found, the new expressions for C_1, C_2, and C_3, and hence the S-parameters, can be obtained by means of (6) and (7).

With the conductivity of the metal denoted by σ and the skin depth by δ $\left(\delta = 1/\sqrt{\pi f \mu \sigma}\right)$, while assuming $e^{j\omega t}$ time dependence, the small parallel component of the electric field can be written as [10]

$$\boldsymbol{E}_{\|} = \frac{1}{\sigma\delta}(1+j)\left(\boldsymbol{n} \times \boldsymbol{H}_{\|}\right). \tag{9}$$

Using orthogonality, an expression for $E_N(r,\phi)$ and hence $\Delta E_N(r,\phi)$ can be found

$$\Delta E_N(r,\phi) = \frac{j}{2\pi}[H_N(R,\phi)\Delta\phi'] \sum_{n=-\infty}^{\infty} \frac{e^{jn(\phi-\phi')}}{D_n} \cdot \left[Z_{\text{eff}} J_n(kr) + \frac{(1-j)}{\sigma\delta}X_n + \frac{(1+j)}{\sigma\delta}Y_n\right] \tag{10}$$

where

$$D_n = J_n'(x) - \frac{\kappa}{\mu} n \frac{J_n(x)}{x}$$
$$X_n = J_n'(kr) - \frac{\kappa}{\mu} n \frac{J_n(kr)}{kr}$$
$$Y_n = \frac{\kappa}{\mu} J_n'(kr) - n\frac{J_n(kr)}{kr}. \tag{11}$$

Denoting the Green's function shown in (1) as G_B after Bosma, the new Green's function in the presence of conductor loss, and assuming $e^{j\omega t}$ time dependence, can be written as

$$G_N = G_{\text{B}} - G_{N1} - G_{N2},$$

where G_{N1} and G_{N2} are the two new terms. After some tedious algebra, both of the new terms in the Green's function can be expressed as a sum of a zeroth-order term and a summation involving positive n's with coefficients attached to $\sin n(\phi - \phi')$ and $\cos n(\phi-\phi')$. The results are

$$G_{N1} = -\frac{(1+j)}{2\pi\sigma\delta}\frac{J_0'(kr)}{J_0'(x)} + \frac{(1-j)}{\pi\sigma\delta}\sum_{n=1}^{\infty} \cdot \frac{(ad-bc)\sin[n(\phi-\phi')] - j(ac-bd)\cos[n(\phi-\phi')]}{(c^2-d^2)}$$

$$G_{N2} = \frac{(1-j)}{2\pi\sigma\delta}\frac{\kappa}{\mu}\frac{J_0'(kr)}{J_0'(x)} + \frac{(1+j)}{\pi\sigma\delta}\sum_{n=1}^{\infty} \cdot \frac{(ed-gc)\sin[n(\phi-\phi')] - j(ec-gd)\cos[n(\phi-\phi')]}{(c^2-d^2)}$$

where

$$a = J_n'(kr)$$
$$b = \frac{\kappa}{\mu} n \frac{J_n(kr)}{kr}$$
$$c = J_n'(x)$$
$$d = \frac{\kappa}{\mu} n \frac{J_n(x)}{x}$$
$$e = \frac{\kappa}{\mu} J_n'(kr)$$
$$g = n\frac{J_n(kr)}{kr}.$$

If the center conductor and the ground plane metals are different, with conductivities and skin depths $\sigma_{1,2}$ and $\delta_{1,2}$, respectively, then the term $\frac{1}{\sigma\delta}$ should be replaced by $\left(\frac{1}{\sigma_1\delta_1} + \frac{1}{\sigma_2\delta_2}\right)$.

The S-parameters developed using this Green's function method are ratios of transimpedances and provide only part of the information for determining conductor losses. By analogy to TEM waves between lossy parallel planes, the attenuation constant is given by

$$\alpha = \frac{8.686 R_S}{\eta h} \text{ dB per unit length},$$

where R_S is surface resistance, η is wave impedance, and h is the height of the ferrite disk used in the analysis (the full height for microstrip and half of the ground plane spacing for stripline). The S-parameters used here are equivalent to the impedance ratio R_S/η in the expression for the attenuation constant. In the circulator, then, it is necessary to use, for total losses,

$$(\alpha l)_{\mathrm{dB}} = 8.686\left(\frac{R_s}{\eta h}\right) l = (S_{mn})_{\mathrm{dB}}\left(\frac{l}{h}\right),$$

where S_{mn} is the scattering parameter from port n to port m, and l is one-third of the ferrite disk circumference $(2\pi R/3)$.

Because of the linear relationship between $R_S = \frac{1}{\sigma\delta}$ and dissipation loss in dB, which can be shown numerically for small losses, the term $\frac{2\pi R}{3h} \times \frac{1}{\sigma\delta}$ can replace $\frac{1}{\sigma\delta}$ in the equations for G_{N1} and G_{N2} above.

A circulator operating at the point of infinite isolation and perfect match is like a two-port circuit and its loss can be calculated in an alternate way as a check on the circulator calculations. A wavelength is $\lambda_g = 4\pi R/3$, since for the case of perfect circulation, the input to output phase shift is 180 degrees. For copper conductors with $R_S = 2.61 \times 10^{-7}\sqrt{f}\ \Omega$ and for $\eta = 377/\sqrt{\epsilon_{r_{eff}}}$, where $\sqrt{\epsilon_{r_{eff}}} = c/(f\lambda_g)$, a simple expression for loss through a length $2\pi R/3$ can be developed as

$$\alpha l(h, f)_{\mathrm{dB}} = 0.002822 \Big/ \left(h_{\mathrm{cm}}\sqrt{f_{\mathrm{GHz}}}\right).$$

This expression gives $\alpha l(0.375, 1.5) = 0.0061$ dB, $\alpha l(0.05, 26) = 0.011$ dB, and $\alpha l(0.0035, 94) = 0.08$ dB; while the corresponding circulator calculations give 0.008, 0.015, and 0.071 dB. This agreement is quite satisfactory, especially since the conductor loss is usually small compared with dielectric and magnetic losses. However, the circulator conductor losses increase with decreasing ferrite thickness. Also, although matching circuits have not been discussed here, their conductor losses are affected indirectly by ferrite thickness, since thinner ferrite means lower impedance level at the circulator, resulting in higher loss for equal bandwidth at the eventual 50 Ω level.

C. Magnetic Losses

In order to include magnetic loss in the calculation of the S-parameters, the quantities μ and κ have to be considered as complex. If $\mu = \mu' - j\mu''$ and $\kappa = \kappa' - j\kappa''$, then the real and imaginary parts of μ and κ can be written as [8]

$$\mu' = 1 + \frac{\omega_m\omega_0\left[\omega_0^2 - \omega^2(1 - \alpha_m^2)\right]}{\left[\omega_0^2 - \omega^2(1 + \alpha_m^2)\right]^2 + 4\omega^2\omega_0^2\alpha_m^2}$$

$$\mu'' = \frac{\omega_m\omega\alpha_m\left[\omega_0^2 + \omega^2(1 + \alpha_m^2)\right]}{\left[\omega_0^2 - \omega^2(1 + \alpha_m^2)\right]^2 + 4\omega^2\omega_0^2\alpha_m^2}$$

$$\kappa' = \frac{-\omega_m\omega\left[\omega_0^2 - \omega^2(1 + \alpha_m^2)\right]}{\left[\omega_0^2 - \omega^2(1 + \alpha_m^2)\right]^2 + 4\omega^2\omega_0^2\alpha_m^2}$$

$$\kappa'' = \frac{-2\omega_0\omega^2\alpha_m\omega_m}{\left[\omega_0^2 - \omega^2(1 + \alpha_m^2)\right]^2 + 4\omega^2\omega_0^2\alpha_m^2},$$

where $\omega_m = -\gamma M \approx -\gamma M_s$ (M_s is the saturation magnetization). The resonance frequency, ω_0, is related to the applied internal magnetic field H_i by $\omega_0 = -\gamma H_i$; the phenomenological damping term α_m is related to the linewidth ΔH by $\alpha_m = -\frac{\Delta H\gamma}{2\omega}$; and γ is the gyromagnetic ratio whose value in rationalized MKS units is -2.21265×10^5 (rad/s)/(ampturns/m).

In the theory from which the equations for μ and κ were derived [8], α_m is considered a constant and the linewidth ΔH is the Ferromagnetic Resonance (FMR) linewidth. Although published measured data are scarce [11], [12], the FMR linewidth of at least some ferrite materials increases with increasing frequency, which would be necessary for α_m to remain constant. In the calculations to be given here, α_m is assumed constant and its value is derived from the material supplier's data at X-Band, using the FMR linewidth. The magnitude of the total losses, using magnetic loss calculated in this way, is in good agreement with reality.

IV. Results

It is difficult to find enough reliable published data on any specific circulator to test the loss theory described here. It would also be very expensive to carry out a unique fabrication and test program of many circulator designs at many different frequencies. But there is a different and perhaps better way to do it, using published information on many state-of-the-art designs over all of the microwave frequency range from 1 to 26.5 GHz. If agreement between measured and calculated losses can be shown over this whole range, using reasonable analysis of published information, then the validity of the theory is confirmed.

The method used here for testing the loss calculations starts with Fig. 4, where the loss results or specifications are summarized for more than 80 state-of-the-art coaxial Y-junction circulators [13]. In Fig. 4, the wideband circulator designs are the "high" loss entries and the narrowband designs have the lowest losses. This is generally because the wideband designs must incorporate more extensive matching circuitry. It is safe to say that the dashed boundary line drawn to connect the lowest loss points in Fig. 4 constitutes something fairly close to the state-of-the-art intrinsic circulator loss at any frequency. Even these values contain some connector losses

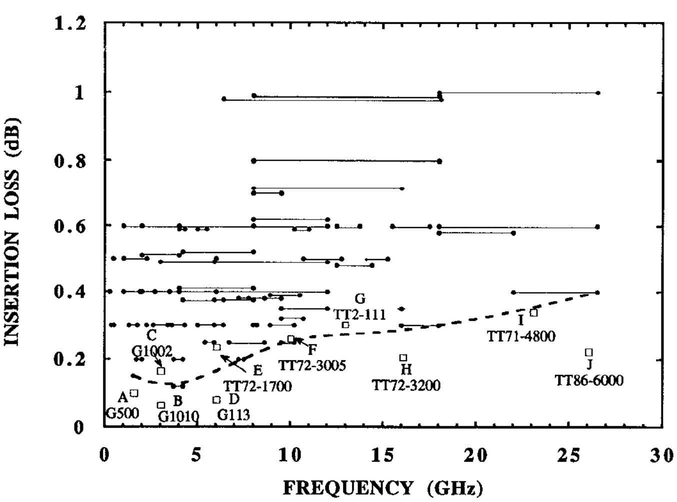


Fig. 4. State-of-the-art coaxial Y-junction circulator insertion loss versus frequency.

so that the actual intrinsic circulator losses must be slightly less than those along the minimum loss connecting line.

Several narrowband stripline circulators were designed by the coarse ground rules in [9], and briefly optimized with the computer program discussed here, using "catalog" low FMR linewidth materials from Trans-Tech. The calculated total losses of these designs (at the highest isolation point) are given by the squares in Fig. 4. The calculated points do not fall on a smooth curve because independent designs were carried out for each frequency. Actual ferrite materials suited for each frequency were used, rather than using hypothetical material parameters which could produce smoother looking results. The material used is given at each point. No matching losses are included in the calculated values, so they are "intrinsic" values. All designs used copper conductors and $\epsilon_r = 10$ for the medium outside the ferrite disk. The magnitudes of these calculated losses are in good agreement with the estimated intrinsic losses of existing circulators, and the slope of the calculated loss versus frequency is like that of existing circulators. Therefore, the theory here appears on solid ground.

It is interesting to see the breakdown of losses into the three types. For instance, at 1.5 GHz using circulator design A, the dielectric loss is 0.002 dB (2%), conductor loss for copper is 0.008 dB (7%), and magnetic loss is 0.111 dB (91%). Then at 26 GHz using circulator J, the dielectric loss is 0.003 dB (1%), conductor loss for copper is 0.015 dB (7%), and magnetic loss is 0.214 dB (92%). An extrapolation to 94 GHz, using TT86-6000 material, gives dielectric loss of 0.11 dB (33%), conductor loss for copper of 0.071 dB (21%), and magnetic loss of 0.154 dB (46%).

A few general observations can be developed from the calculations completed, using modern ferrite/garnet materials.

- There is a strong correlation between low intrinsic losses and narrow FMR linewidth, as shown by calculated points $B(\Delta H = 40$ Oe$)$ and $C(\Delta H = 100$ Oe$)$ and by points $D(\Delta H = 45$ Oe$)$ and $E(\Delta H = 150$ Oe$)$.
- The intrinsic losses of coaxial Y-junction circulators are small, and magnetic loss is usually the largest contributor.
- Wideband circulator losses are dominated by matching circuit losses, not by the intrinsic circulator losses. In Fig. 4 of [5], the measured data on a broadband (3–10 GHz) stripline circulator show a bit less than 1 dB of insertion loss at 7 GHz, the approximate frequency of the isolation maximum, most of which is dissipation loss. At that frequency, the calculated total intrinsic loss of the circulator disk is only 0.08 dB, while the calculated loss of two passes through the 8–50 Ω matching tapers is 0.9 dB for copper conductors and alumina dielectric. These tapers are very long for experimental purposes and do not necessarily provide the lowest loss achievable.
- Intrinsic conductor loss is usually negligible, perhaps even at mm-wave frequencies, unless the circulator is extraordinarily thin.
- Dielectric losses are usually negligible at lower frequencies, but may dominate at mm-waves. This observation is supported by the results of a waveguide example in [14], where the major importance of material loss tangent and resultant dielectric loss at 95 GHz is discussed.

An example of the effect of loss on the three-port characteristics is shown in Figs. 5 and 6. The circulator design is the same as that which produced the G data point in Fig. 4. The insertion loss result in Fig. 5 increases as expected. But the return loss and the isolation are seen to increase or to decrease, relative to the lossless values, depending on frequency. The reference characteristic impedance for the curves of Fig. 5 is 14.5 Ω. The effect on input impedance is shown in Fig. 6. The input resistance increases a very small amount, and the reactance change may be interpreted as a slight shift downward in center frequency.

The work done here is specifically for a circular ferrite resonator shape. Koshiba and Suzuki [15], using a numerical approach, calculated the magnetic dissipation loss for TT1-109 circulator resonators of three different shapes in a WR-90 waveguide junction. Their loss results at the frequency of the isolation peak, when that frequency is 10 GHz, vary from about 0.05 dB for a depressed-side triangular resonator to about 0.08 dB for a circular resonator. This suggests that when extremely low total dissipation loss is needed, there may be

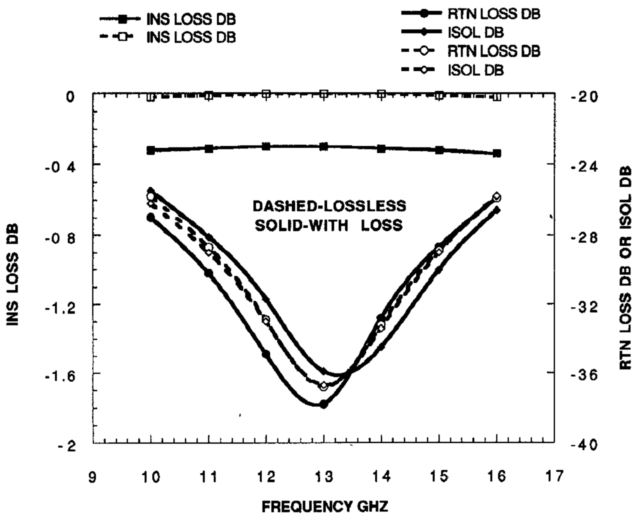


Fig. 5. Circulator three-port performance versus loss.

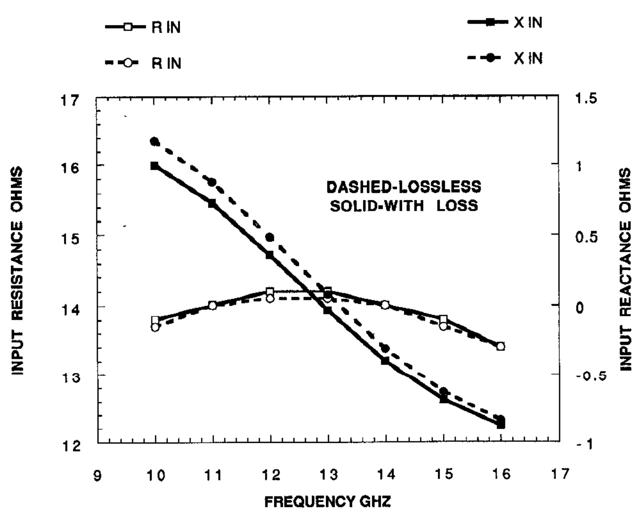


Fig. 6. Circulator input impedance versus loss.

some loss advantage to shapes other than circular; but the applications must be narrowband so that intrinsic circulator losses are not overwhelmed by matching circuit losses.

V. Summary

The main objective here has been to produce a theory to calculate the intrinsic performance of Y-junction circulators with losses, from physical dimensions and magnetic properties of the ferrite material and embedding structure. This has been done, with full 3–port S-parameter calculations in the presence of dielectric, conductor, and magnetic losses. Excellent agreement with other authors' lossless calculations has been shown, and reasonable loss predictions are made. This gives confidence that the theory can be used for further work on circulator improvements and on the general understanding of dissipation losses.

Acknowledgment

Technical discussions with D. C. Webb and A. K. Ganguly of NRL are gratefully acknowledged.

References

[1] R. E. Neidert, "Computer aided approach to the design of Y-junction stripline and microstrip ferrite circulators," Naval Res. Lab. Rep. NRL/FR/6851-92-9381, May 1992. (Copies may be ordered by requesting AD-A251337 from: National Technical Information Service, 5285 Port Royal Road, Springfield, VA 22161, USA. Phone: (703) 487-4650, Fax: (703) 321-8547.)

[2] H. Bosma, "On stripline Y-circulation at UHF," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-12, pp. 61–72, Jan. 1964.

[3] C. E. Fay and R. L. Comstock, "Operation of the ferrite junction circulator," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-13, pp. 15–27, Jan. 1965.

[4] Y. S. Wu and F. J. Rosenbaum, "Wide-band operation of microstrip circulators," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-22, pp. 849–856, Oct. 1974.

[5] E. Schlömann and R. E. Blight, "Broad-band stripline circulators based on YIG and Li-ferrite single crystals," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-34, pp. 1394–1400, Dec. 1986.

[6] J. W. Simon, "Broadband strip-transmission line Y-junction circulators," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-13, pp. 335–345, May 1965.

[7] S. J. Salay and H. J. Peppiatt, "Input impedance behavior of stripline circulator," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-19, pp. 109–110, Jan. 1971.

[8] R. F. Soohoo, *Theory and Application of Ferrites.* Englewood Cliffs, NJ: Prentice-Hall, 1960.

[9] D. K. Linkhart, *Microwave Circulator Design.* Norwood, MA: Artech House, 1989.

[10] J. D. Jackson, *Classical Electrodynamics,* 2nd ed. New York: Wiley, 1975, p. 338.

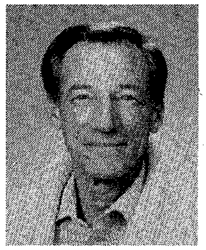
[11] W. H. von Aulock, *Handbook of Microwave Ferrite Materials.* New York: Academic Press, 1965, p. 367.

[12] P. E. Tannenwald, "Ferromagnetic resonance in manganese ferrite single crystals," *Phys. Rev.,* vol. 100, pp. 1713–1719, 1955.

[13] G. P. Rodrigue, "Circulators from 1 to 100 GHz," *Microwave J.,* 1989 State of the Art Reference, p. 115ff, 1989.

[14] G. R. Harrison, S. B. Thompson, J. T. Vaughn, and G. P. Rodrigue, "Impact of dielectric loss tangent on the performance of millimeter wave ferrite circulators," in *MTT-S Int. Microwave Symp. Dig.,* vol. II, June 1987, pp. 989–991.

[15] M. Koshiba and M. Suzuki, "Finite-element analysis of H-plane waveguide junction with arbitrarily shaped ferrite post," *IEEE Trans. Microwave Theory Tech.,* vol. MTT-34, pp. 103–109, Jan. 1986.



Robert E. Neidert (S'56–M'70) received the B.E. (E.E.) degree from Vanderbilt University, Nashville, TN, in 1959, and has done graduate work at the University of Florida, St. Petersburg.

From 1959 to 1962 he was with the Sperry Microwave Electronics Company, Clearwater, FL, where he was engaged in the development of microwave components for radar systems. From 1962 to 1969 he served as Senior Design Engineer and Project Leader at the General Electric Company, Communications Products Department, Lynchburg, VA, in the design and development of microwave components and solid-state sources for TV and multiplex telephone radio relay equipment. From 1969 to 1972 he was a principal Engineer at Radiation Systems, Inc., McLean, VA, where his work was in antenna and antenna feed network design. Since 1972 he has been involved in research on microwave and millimeter-wave devices and circuits at the Naval Research Laboratory, Washington, DC. He has authored numerous papers in the fields of communications systems components, microwave integrated circuits, computer-aided microwave circuit design, bipolar and FET amplifier design, FET modeling, millimeter-wave circuits, and vacuum microelectronics.

Mr. Neidert is a member of Tau Beta Pi.

Purobi M. Phillips (M'91) was born in India on April 19, 1961. She received the B.Sc. degree in physics from Presidency College, Calcutta, India, in 1983, and the A.M. and Ph.D. degrees in physics from Dartmouth College in 1985 and 1987, respectively.

Since 1987 she has been a Research Scientist with Science Applications International Corporation's Plasma Physics division where she is assigned to work in the Electronics Science and Technology Division of the Naval Research Laboratory. Her current research area is vacuum microelectronics.